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APPLICATION

FOR

UNITED STATES LETTERS PATENT

TITLE: DIGITAL RECEIVER HAVING ADAPTIVE CARRIER RECOVERY CIRCUIT

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DIGITAL RECEIVER HAVING ADAPTIVE CARRIER RECOVERY CIRCUIT

FIELD OF THE INVENTION

[0001] The present invention relates generally to digital signal receivers and more particularly to a receiver including an adaptive carrier recovery circuit and methods for such receivers.

BACKGROUND OF THE INVENTION

[0002] In many modern digital transmission systems, digital transmitters modulate digital information onto radio frequency carrier signals. The modulated information signals are propagated to digital signal receivers where the digital signals may be demodulated and decoded.

[0003] For example, quadrature amplitude modulation (QAM), as detailed in International Telecommunications Union (ITU) recommendation ITU-T J.83B/SCTE DVS-031/DOCSIS, and vestigial side-band (VSB) modulation, as detailed in the Advanced Television Standard Committee (ATSC) standard A53/B, are two common digital modulation techniques that are used for cable and terrestrial digital television broadcast.

[0004] Invariably, propagation media for the signals introduce distortions in signals received at the receiver. Distortions may, for example, be introduced as a result of noise, phase shifts, signal attenuation, and multi-path interference. Imperfections at the transmitter may similarly introduce distortions in the transmitted signal. Ultimately, these distortions may manifest themselves as increased bit errors in the decoded digital signal at the receiver.

[0005] To reduce the overall bit error rate in the decoded signals, introduced distortions may be compensated in numerous ways. For

example, to mitigate the effects of additive noise, the digital signals may include forward error correcting (FEC) codes. One or more equalizers may be used at the receiver to compensate for phase and amplitude distortions introduced into the transmitted signal by the propagation medium, such as multipath distortion.

[0006] For a terrestrial radio-frequency (RF) propagation medium, the characteristic of the propagation medium varies in dependence on a number of factors, including the location of the receiver. Accordingly, any equalizer at the receiver is typically adaptive and allows its parameters to be matched to the propagation medium at the receiver. The equalizer should also be able to compensate for variations in the propagation medium over time.

[0007] An example digital receiver including an adaptive equalizer is, for example, more particularly detailed in U.S. Patent No. **6,418,164**, the contents of which are hereby incorporated by reference. As disclosed, a conventional digital receiver may include a feed forward equalizer (FFE), a decision feedback equalizer (DFE), and a carrier recovery circuit, each designed to compensate for distortion in the received signal.

[0008] The FFE and DFE minimize multipath inter-symbol interference. Specifically, the FFE mitigates the effect of delayed echo multipath distortion ("pre-cursor"), while the DFE minimizes leading ("post cursor") multipath interference. Error metrics between the demodulated signal and the reconstructed signal are used to provide closed loop adjustment of equalizer parameters. The carrier recovery circuit compensates for shifts in phase and/or of the carrier frequency of the broadcasted signal. The aim is to adjust simultaneously FFE and DFE equalizer and carrier recovery circuit parameters to compensate for the propagation medium distortion. A blind equalization technique using for example, the Constant Modulus Algorithm (CMA), may be used to adjust equalizer parameters.

[0009] Although very effective for channels having static characteristics, the disclosed receiver may not be able to track fast changes in the carrier frequency or phase. This is particularly acute when the receiver is used to

receive VSB modulated signals, which are highly carrier dependent. Indeed, for VSB signals phase distortion affects real and complex values of the VSB constellation differently. Furthermore, a CMA will not be phase independent when applied to a signal modulated with a VSB constellation, as it is with a QAM modulation for example

[0010] Accordingly, there is clearly a need for a receiver including a carrier recovery circuit that can track quick changes in carrier phase/ frequency without otherwise impacting the overall performance of the receiver.

SUMMARY OF THE INVENTION

[0011] A digital receiver includes an adaptive fine carrier recovery circuit that compensates for deviations in the carrier frequency or phase. The fine carrier recovery circuit de-rotates a signal including phase errors. Estimations of phase errors are filtered using a filter whose gain and bandwidth are adjusted adaptively. This allows the carrier recovery circuit to track phase/ frequency offset without introducing significant jitter. In one embodiment, the receiver includes a DFE, and the adaptive carrier recovery circuit mitigates instability that might be associated with the DFE.

[0012] In accordance with an aspect of the present invention, there is provided a digital receiver for demodulating a digital signal modulated onto a carrier, includes a tuner; an analog to digital converter, for digitizing a channel tuned by the tuner to provide a digitized channel; a coarse carrier recovery circuit for extracting the signal from the digitized channel at near baseband; a feed forward equalizer receiving the signal at near baseband and outputting a feed forward equalized signal; a fine carrier recovery circuit for phase shifting the feed forward equalized signal by a phase correction angle to adjust for remaining offsets in phase and frequency of the feed forward equalized signal attributable to phase and frequency offsets in the carrier. The fine carrier recovery circuit includes a filter for filtering an estimate of a phase error in the carrier to control the phase correction angle. At least one filter parameter of the filter varies adaptively with the phase error.

[0013] In accordance with another aspect of the present invention there is provided, in a digital receiver for receiving a signal modulated onto a carrier, a method includes reducing multi-path interference in the signal, by filtering the signal through a feed-forward equalizer to produce a feed forward-equalized signal; determining an estimate of a remaining phase error in the carrier; filtering the estimate through a filter having at least one adjustable filter parameter to produce a phase correction signal; varying the adjustable parameter with the estimate of phase error; and multiplying the feed forward equalized signal by the phase correction signal to de-rotate the feed forward equalized signal.

[0014] In accordance with yet another aspect of the present invention there is provided a digital receiver for demodulating a digital signal modulated onto a carrier, to produce a demodulated digital signal. The receiver includes a de-rotator for phase shifting an equalized version of the digital signal by a phase correction angle to adjust for remaining offsets in phase and frequency of the equalized version of the digital signal attributable to phase and frequency offsets in the carrier, and a filter in communication with the de-rotator, for filtering an estimate of a phase error in the demodulated digital signal to control the phase correction angle. At least one filter parameter of the filter varies adaptively with the phase error.

[0015] Other aspects and features of the present invention will become apparent to those of ordinary skill in the art upon review of the following description of specific embodiments of the invention in conjunction with the accompanying figures.

BRIEF DESCRIPTION OF THE DRAWINGS

[0016] In the figures which illustrate by way of example only, embodiments of the present invention,

[0017] **FIG. 1** is a simplified schematic block diagram of a VSB receiver exemplary of an embodiment of the present invention;

[0018] **FIG. 2** is a simplified schematic block diagram of a carrier loop of the receiver of **FIG. 1**;

[0019] **FIG. 3A-3E**, **4A-4B** and **5A-5B** are simplified block diagrams of portions of a phase error detector of the carrier loop of **FIG. 2**;

[0020] **FIG. 6** is a simplified schematic block diagram of an adaptive loop filter of the carrier loop of **FIG. 2**;

[0021] **FIG. 7** is a simplified schematic block diagram of an adaptive gain circuit of the carrier loop of **FIG. 2**;

[0022] **FIG. 8** is a simplified schematic block diagram of a low pass filter of the carrier loop of **FIG. 2**; and

[0023] **FIG. 9** is a simplified schematic block diagram of a QAM receiver exemplary of an embodiment of the present invention.

DETAILED DESCRIPTION

[0024] **FIG. 1** illustrates a VSB receiver **10**, including a carrier recovery circuit **29**, exemplary of an embodiment of the present invention. VSB receiver **10** may for example be used to receive a digital television signal, as specified by the Advanced Television Systems Committee (ATSC) in the standard A53/B. Receiver **10** may be formed as an integrated circuit using conventional application specific integrated circuit manufacturing techniques.

[0025] Receiver **10** is interconnected to an antenna **12**, providing a radio frequency signal to a tuner **14**. Tuner **14**, in turn selects a tuned ATSC digital 6Mhz wide channel of interest at an assumed frequency in a conventional manner. Tuner **14** translates the tuned channel to an intermediate frequency (IF) so that all channels can be processed at a single unique frequency by the receiver front end **16**. Front end **16** scales the received channel by automatic gain control amplifier **17** to allow the channel to be digitized by analog to digital (A/D) converter **18**. A/D converter **18** samples the analog channel at the IF frequency and provides a digital signal corresponding to the digitized channel at second IF frequency. The digitized channel is passed to timing

recovery block **20** where timing information is extracted, in a manner understood by those of ordinary skill. The timing information establishes the clock rate and phase at which symbols forming the signal of interest modulated on the carrier are detected. The digitized channel is optionally amplified by automatic gain control (AGC) block **21** to ensure the magnitude samples of channels are optimal for processing by the remainder of receiver **10**.

[0026] The digitized channel is converted to near baseband through a frequency shift based on the assumed carrier frequency, at coarse carrier recovery circuit **22**. Coarse carrier recovery circuit **22** may be formed in any number of ways appreciated by those of ordinary skill. Coarse carrier recovery circuit **22** may for example be formed as a frequency and phase-locked loop adapted to detect and lock to a low level pilot signal contained in the received ATSC VSB channel. In any event, coarse carrier recover circuit **22** extracts a near baseband digital signal that is provided to feed-forward equalizer circuit **24**. The signal provided is complex and thus includes two parallel data streams.

[0027] As will be appreciated, the description of front end **16** is exemplary only. A suitable front end **16** could be formed in many ways. For example, the order of coarse carrier recovery circuit **22** and timing recovery block **20** could be reversed. A person of ordinary skill will recognize many other design alternatives.

[0028] Input to detection circuit **24** includes the sequence of VSB data symbols corrupted by noise, phase shifts, and inter-symbol interference. In particular, the signal includes errors due to variations in carrier phase/frequency caused by carrier phase noise, not compensated for by carrier recovery circuit **22**.

[0029] Feed-forward equalizer **26** in conjunction with decision feedback equalizer **36** corrects signal degradation resulting from inter-symbol interference by filtering the input signal to detection circuit **24** and the output of slicer **38** with digital filters having variable coefficients. Exemplary FFE **26**

is a linear adaptive filter having complex coefficients. FFE **26** may be formed as a finite impulse response (FIR) filter having complex coefficients, as detailed in "Current approaches to blind decision feedback equalization", by R.A. Casas, T.J. Endres, A. Touzni, C.R. Johnson, Jr. in Signal Processing Advances in Wireless and Mobile Communications – Trends in Channel Estimation and Equalization. Vol 1. Chap 11, pp. 367-415, 2000.

[0030] FFE **26** takes complex valued data from coarse carrier recovery circuit **22**. By using complex values instead of a real values filter, the equalization performance of the FFE **26** is improved. However, errors due to variations in carrier phase/frequency caused by carrier phase noise are still present after FFE **26**.

[0031] As such, the carrier phase/frequency error is corrected using a carrier recovery circuit **29**. Such correction is referred to as "de-rotation". Example carrier recovery circuit **29** includes a carrier loop **30** and a mixer/multiplier **28**. Carrier recovery circuit **29** multiplies the signal output at FFE **26** by a sinusoidal signal produced by carrier loop **30** to produce a complex output, derot(n), for each symbol modulated onto the carrier. The multiplication rotates the modulated signal points about the origin of the signal plane. Multiplication by the sinusoidal signal compensates for any phase / frequency deviation of the carrier signal, as detailed below. At steady state, carrier recovery circuit **29** tracks the frequency and phase of the signal. Common approaches for signal de-rotation are, for example, detailed in, Meyr et al., Digital Communication Receivers, John Wiley & Sons, 1998.

[0032] As receiver **10** is a VSB receiver, the signal information is carried in the real component of the modulated symbols. The real component of the output of carrier recovery circuit **29**, derot(n), extracted at block **32** and provided to slicer **38**. Slicer **38** produces a demodulated, quantized signal S(n) of the real value of the complex signal constellation.

[0033] A DFE **36** filters a version of the quantized output signal S(n), and subtracts this version to the output derot(n) of carrier recovery circuit **29** to form y(n), which is a real valued estimation of the transmitted complex

symbol, provided to the input of slicer **38**.

[0034] DFE **36** in conjunction with FFE **26** mitigates the effects of multipath interference. DFE **36**, like FFE **26** may be formed as an adaptive FIR filter, as detailed in "Current approaches to blind decision feedback equalization", *supra*.

[0035] DFE **36** is a linear adaptive filter having real values for both data and coefficients. DFE **36** mitigates post-cursor multi-path interference by feeding back symbol estimates from the slicer **38**. However, DFE **36** easily becomes unstable especially when the output of the FFE has not been properly corrected in phase. Indeed, in this case, even a perfect inter-symbol correction provided by FFE **26** and DFE **36** may not compensate for a phase distortion of the output of FFE filter. This induces a compression of the soft estimation y(n) that may trigger an erroneous decision of the slicer **38**. That is, an erroneous decision of slicer **38** will be propagated in the DFE **36** and may lead to less effective inter-symbol equalization by DFE **36**. This is particularly acute for VSB modulated data since DFE **36** operates on real data, and carrier information is lost. In such circumstances DFE **36** cannot carry any phase information as FFE **36** does.

[0036] For VSB modulation such as in ATSC A53/B, for each modulated symbol, slicer **38** decodes y(n) to produce a 8-level or a 16-level discrete VSB symbol S(n) corresponding to the real value contribution of the complex 8/16-VSB constellation broadcasted by the transmitter. Each VSB symbol S(n) corresponds to an allowable VSB constellation. If slicer **38** is adapted to decode 8-VSB signals, each data symbol takes on one of eight relative values: –**7**, -**5**, -**3**, -**1**, **1**, **3**, **5**, and **7**. For 16-VSB signals, each data symbol takes on one of sixteen relative values: **-15**, **-13**, **-11**, **-9**, **-7**, **-5**, **-3**, **-1**, **1**, **3**, **5**, **7, 9, 11, 13, 15**. As will be appreciated, slicer **38** may be formed as a trellis decoder to optimally decode the VSB signals. As will become apparent, the signal S(n), the input y(n) to slicer **38** and the output of block **34** are used to estimate the carrier phase error by computing appropriate error metrics. The metrics are calculated by carrier loop **30**, as detailed below. A similarly calculated error metric is also provided to FFE **26** and DFE **36** to adjust their

filter parameters. Operating parameters of DFE **36** and FFE **26** may be adjusted using a simplified decision directed constant modulus algorithm and the approximated phase error, as detailed in U.S. Patent No. **6,337,878**, the contents of which are hereby incorporated by reference.

[0037] As will be appreciated, for ATSC signals, the sequence of symbols y(n) are organized as frames. FEC decoder **40** receives the input y(n) to slicer **38**, and further corrects errors in the demodulated digital signals using FEC. The corrected decoded data is provided to a decoder **42**, which may be a MPEG-2 decoder, or the like, to provide audio and video data of interest.

[0038] **FIG. 2** further illustrates an example carrier loop **30**, exemplary of an embodiment of the present invention. As illustrated, carrier loop **30** includes a multi-mode carrier phase error detector **50**, a loop filter **52**, an adaptive gain circuit **54**, a low pass filter **56** to filter the output of adaptive gain circuit **54**, a sine/cosine generator **58**, and a mode control block **60**.

[0039] Mode control block **60** assumes a number of states based on the state of receiver **10** (e.g. initial signal acquisition mode, tracking mode, etc.) and the quality of the received signal, and generates control signals to select the appropriate phase error detector **50**, and to loop filter **52** controlling one of multiple modes of operation of each of detector **50** and loop filter **52**, as detailed below.

[0040] Example carrier phase error detector **50** computes the phase error in the carrier from three inputs: input to slicer **38** (y(n) – **FIG. 1**), carrier corrected output of FFE **26** (real component of the complex valued derot(n)); and output of slicer **38** (S(n)). Using these inputs, detector **50** estimates the error that is caused by offset between carrier frequency and phase at the receiver and transmitter. Phase error in the carrier may be estimated in a number of ways. For example, phase error may be estimated by calculating the phase difference between the demodulated signal at the input and output of slicer **38**.

[0041] Example carrier phase error detector **50** includes seven phase error detector calculation blocks illustrated in **FIGS. 3A-3E**, **FIGS. 4A-4B** to

calculate metrics to estimate the error in the imaginary (errorImag(n)) and real part (errorReal(n)) of the demodulated signal. These blocks are labelled DD_CL (block **62a** – **FIG. 3A**), DD_CL_ALT (block **62b** - **FIG. 3B**), MSE_SINGLE_CL_I (block **62a** – **FIG. 3C**), MSE_SINGLE_CL_Q (block **62d** – **FIG. 3D**), QPSK_CL (block **62e** - **FIG. 3E** and **64a** – **FIG. 4A**) and other (block **64b** – **FIG. 4B**).

[0042] To appreciate the operation of blocks **62a-62e** and **64a-64e**, it may be noted that if y(n) and S(n) were complex signals, and if $y_i(n)$ and $y_q(n)$ denote respectively the real value and the imaginary contribution of the complex value y(n), and if $S_i(n)$ and $S_q(n)$ denote respectively the real value and the imaginary contribution of the complex value S(n), errorImag(n) may be determined as:

[0043] $(y_i(n)*S_q(n)-y_q(n)S_i(n))*$errorNormImag (block **62a**, **FIG. 3A**);

[0044] sign(derotq(n))*$(y_i(n) - S_i(n))$ *errorNormImag (block **62b**, **FIG. 3B**); where sign(derotq(n)) extracts the sign of derot(n)

[0045] $derot_i(n) * S_q(n) *$ errorNormImag (block **62c**, **FIG. 3C**);

[0046] $-derot_q(n) * S_i(n) *$ errorNormImag (block **62d**, **FIG. 3D**); or

[0047] $y_i(n)*qpsk_q(n)-y_q(n)*qpks_i(n)$ (block **62e**, **FIG. 3E**);

[0048] Similarly, errorReal(n) may be determined as,

[0049] $(y_i(n)*qpsk_i(n)+y_q(n)*qpsk_q(n))$ (block 64**a**, **FIG. 4A**); or

[0050] $(S_i(n)*y_i(n)+S_q(n)*y_q(n))*$errorNormImag, (block 64**b**, **FIG. 4B**);

[0051] where errorNormImag(n), and errorRealnorm(n) may be derived from the magnitudes of $S_i(n)$ and $S_q(n)$. These normalization factors normalize the amplitude of the different metrics and thus the phase error estimation resulting form the metrics. The normalization factors are formed from $S_i(n)$ and $S_q(n)$ typically using a look-up table by block **100a**, of phase error detector **50** illustrated in **FIG. 5B**.

[0052] errorNormImag may be calculated by block **100a** as,

[0053] errorNormImag=$1/(2\pi^*(| S_i(n) | + | S_q(n)|))$, where $| S_i(n)|$ and $| S_q(n) |$ denote respectively the magnitude of $S_i(n)$ and $S_q(n)$.

[0054] For the error metric calculated by block **62b**, the normalization factor can be simplified to errorNormImag=$1/2\pi^*| S_i(n) |$, since $S_q(n)$ can be considered zero. For metrics calculated by blocks **62c** and **62d**, errorNormImag=$1/2\pi^*(| S_q(n) |)$ since $S_i(n)$ can be interpreted as a null component. Usually a minimal normalization value is applied when both components are close to zeros.

[0055] $qpsk_i(n)$ and $qpsk_q(n)$ are quadrature phase keyed approximations of $y_i(n)$ and $y_q(n)$, respectively. These are formed from $y_i(n)$ and $y_q(n)$ using a look-up table by block **100b**, of phase error detector **50** illustrated in **FIG. 5B**.

[0056] Specifically, block **100b** forms $qpsk_i(n)$ and $qpsk_q(n)$ as follows:

[0057] $qpsk_i(n) = 1/(16^*\pi)$ if $y_i(n)>0$ and

[0058] $qpsk_i(n) = -1/(16^*\pi)$ if $y_i(n)<0$,

[0059] $qpsk_q(n) = 1/(16^*\pi)$ if $y_q(n)>0$ and

[0060] $qpsk_q(n) = -1/(16^*\pi)$ if $y_q(n)<0$

[0061] No normalization is required for the qpsk calculations as these are already normalized.

[0062] For VSB modulation, S(n) and y(n) are real. The following may be used to assess phase error using blocks **62a-62e** and **64a** and **64b** and **100a** and **100b**. The input $S_i(n)$ and $S_q(n)$ and the output $y_i(n)$ and $y_q(n)$ are interpreted as follow:

[0063] $S_i(n)=S(n)$, $S_q(n)=S(n-1)$

[0064] $y_i(n)=y(n)$, $y_q(n)=y(n-1)$

[0065] The inputs for the exemplary QPSK phase detector are:

[0066] $qpsk_i(n) = qpsk(n)$, $qpsk_q(n) = qpsk(n-1)$

[0067] Phase error detector **50** thus generates two error signals; errorImag(n) and errorReal(n) using one of blocks **62a**-**62e** for errorImag(n), and block **64a** for errorReal(n) when **62e** is used for errorImag(n) and block **64b** for errorReal(n), for all other phase detector functions are used.

[0068] Typically for QAM modulation the errorImag(n) phase detectors **62a**, **62e** and the errorReal(n) error signals **64a** and **64b** are the only signals used to compute the phase/frequency estimation.

[0069] Specifically, mode control block **60** provides a signal to select one of three modes to phase error detector **50** reflective of whether the quality of the received signal demonstrates high reliability; low reliability; or intermediate reliability. The quality of the received signal may be assessed in any number of ways by mode control block **60**, including measuring errors in reference/training portions of the received signal, measuring the signal to noise ratio, measuring the evolutions of the equalizer coefficient over time, measuring estimates of the signal spectrum over time, measuring variations of the signal in time, etc. or any suitable combinations of such measurements. Mode control block **60** may accordingly additionally be provided with signals allowing such measurements.

[0070] Typically the results of block **62a** or **62b** (**FIGS. 3A**, **3B**), DD_CL or DD_CL_ALT and block **64b** (**FIG. 4B**), may be used when the signal is reliable and the receiver **10** is tracking the demodulated signal. In one embodiment, the reliability of the signal may be estimated by monitoring the variance of symbols y(n) over time in conjunction with a measure of variance over time of one or multiple coefficients of the DFE **36** and FFE **26**. The results of block **62e** (**FIG. 3E**) and block **64a** (**FIG. 4B** - QPSK_CL) may be used for a low reliability signal, exhibited by high variance coefficients. Typically, these phase detectors will be used during a cold start initialisation of the receiver. Practically, the output of blocks **62e** and **64a** (QPSK_CL detector) may, for example be used for ATSC signals after frames are detected. The results of block **62c** or **62d** (**FIGS. 3C, 3D** -

MSE_SINGLE_CL_I, MSE_SINGLE_CL_Q) and block **64b** (**FIG. 4B** - Other) may be used between these extreme signal conditions, such as for example, in the presence of an atypical event, which may lead to the loss of signal equalization. Atypical events include burst noise detection, abrupt change is signal gain, high jitter in the output of blocks **62a** or **62b** or abrupt changes in the adaptation of the coefficients of the FFE **26** or DFE **36**.

[0071] In any event, the estimation of phase error (using errorImag(n)) is filtered by loop filter **52** to provide a phase correction angle estimate $\phi(n)$. Sine/cosine signal generator **58** converts phase correction angle $\phi(n)$ to generate a signal which is mixed by multiplier **28** with the output of FFE **26** (**FIG. 1**) to de-rotate the output of FFE **26** to correct for carrier phase and frequency offsets. In the disclosed embodiment, sine/cosine generator **58** provides two output values corresponding to the sine and cosine of $\phi(n)$ derived from the phase error estimate errorImag(n), calculated by phase error detector **50** and loop filter **52**. Sine/cosine values may be calculated by a suitable look-up table. Mixer **28** (**FIG. 1**) multiplies the complex signal derot(n) by the complex quantity cos$\phi(n)$ + jsin$\phi(n)$ (i.e. $e^{j\phi(n)}$).

[0072] Advantageously, loop filter **52** takes as adaptive gain inputs, the output of an adaptive gain circuit **54** and an associated low pass filter **56**. Adaptive gain circuit **54** takes errorImag(n) and errorReal(n) and adapts the gain of loop filter **52** to optimize the performance of carrier recovery circuit **29**.

[0073] **FIG. 6** illustrates loop filter **52** of carrier loop **30**. Example filter **52** is formed as second order phase locked loop (PLL) in order to track both the phase and frequency offset in the carrier. Loop filter **52** may be operated in one of three modes: a non-adaptive mode of operation and two adaptive modes. Again, the mode of operation of loop filter **52** is controlled by mode control block **60**, in dependence on the operating state of receiver **10**, and the quality of the received signal. The non-adaptive mode may be used during receiver initialization; the first adaptive mode may be used during signal tracking (high quality signal reception); the second adaptive mode may be used for low and intermediate quality signals.

[0074] Typically, the mode of operation for the phase error detector **50** is selected in tandem with a specific phase error detector. For instance the phase error detector of block **62e** (QPSK_CL phase detector) will be used with a fixed gain algorithm, whereas the phase error detector of block **62a** or block **62b** (DD_CL; DD_CL_ALT phase detectors) will be used with an adaptive phase detector. In the depicted embodiment, and as shown in **FIG. 6**, all adaptive gains are computed in parallel, but are used by filter **52** only when needed. This mode of operation allows seamless transitions between fixed and variable gains.

[0075] In its non-adaptive mode, gain of loop filter **52** is programmable through three programmable registers k**0**, k**1**, k**2**. Suitable values of k**0**, k**1** and k**2** for the non-adaptive mode of operation may be hard coded, or provided by a controller (not shown) controlling overall operation of receiver **10**.

[0076] As illustrated, a multiplier **70** multiplies errorImag(n) by a scale factor k**1**. errorImag(n) is similarly multiplied by k**2** at multiplier **68**. Delayed versions of multiplied signals are fed-back through delay blocks **78** and **80**, and added by adders **74** and **76**. A further delayed scaled version of the delayed multiplied signal, delayed by delay block **77** and scaled by the value of k**0** by multiplier **79** is fed back to adder **74**. As a result, the general equations for the loop filter **52,** in its non-adaptive mode may be described as follows.

$$\theta(n) = \theta(n-1) + k2error\,\mathrm{Im}ag(n-1)$$
$$\phi(n) = (1-k0)\times\phi(n-1) + k1\times error\,\mathrm{Im}ag(n-1) + \theta(n)$$

where errorImag(n) is the output of phase error detector **50** for symbol S(n). $\theta(n)$ is now the frequency offset estimate, and $\phi(n)$ is phase offset estimate. As will be appreciated, filter **52**, in its non-adaptive mode, is second order.

[0077] As will be appreciated, the above mathematical descriptions of the loop filter are in time domain. k**0** increases the range of phase errors to which loop filter **52** can lock. k**0** induces a low pass filtering of the phase estimation

which contributes to reduce impulsive phase noise that may be triggered in the VSB receiver by abrupt changes in the equalizer coefficients.

[0078] Loop filter **52** may be used in its non-adaptive mode at start-up (as controlled by mode control block **60**). In the event the errorImag(n) exceeds a threshold, as may for example be caused by noise, errorImag(n) is clipped at block **66**. In the event the errorImag(n) does exceed the threshold, a value of zero (0) is provided by block **64**, ensuring stability of loop **30**.

[0079] In its adaptive modes **k2** is not used (i.e. k2 is effectively set to zero) and k1adapt or k1adaptavg is used in place of **k1**. As a consequence loop filter **52** calculates

$$\theta(n) = \theta(n-1)$$
$$\phi(n) = (1-k0)x\phi(n-1) + k1adapt \times error\,\mathrm{Im}\,ag(n-1) + \theta(n)$$
$$[or\phi(n) = (1-k0)x\phi(n-1) + k1adaptavg \times error\,\mathrm{Im}\,ag(n-1) + \theta(n)]$$

Of course filters **52** could be made adaptive in many other ways. In this mode, the frequency offset $\theta(n)$ is assumed to be correct and a constant over time. Unlike in its non-adaptive mode, exemplary filter **52** is first order in its adaptive mode. The phase offset $\phi(n)$ is calculated with a first order loop filter for which the gain is adaptive (i.e. the gain is a function of the past observation). The adaptive gain allows the carrier loop to adapt to abrupt variations of phase/frequency.

[0080] A person skilled in the art will now recognize that exemplary loop filter **52** calculates over time $\theta(n)$ that minimizes an error comparable to the error detected by phase detector **50**, operating in its selected mode using gradient stochastic optimisation. The gradient stochastic optimization uses variable gain k1adapt computed by block **54**. Variable gain k1adapt is calculated and adjusted adaptively as described below to track the variation of the channel over time which allows atrade-off between a fast change in the channel and an accurate estimate of the phase. As will become apparent, mathematically in an exemplary embodiment an optimal gain k1adapt is calculated to optimize a cost function. For example, in the case of the single-

axis constant modulus, cost $\sum_{n}(\mathrm{Real}(y(n)*\exp(j\theta(n))^{2}-1)$ is optimised with respect $\theta(n)$ and the gain k1 under the constraint that the gradient estimate of $\theta(n)$ can be expressed in the form $\theta(n)=\theta(n-1)$ + k1 Δ, where Δ is the derivative of the constant modulus cost function with respect to θ. To estimate k1adapt, another gradient optimisation with respect k1 is used.

[0081] **FIG. 7** schematically illustrates the adaptive gain controller **54** of carrier loop **30**. Adaptive gain controller **54** calculates k1adapt(n) from past values of the phase error as estimated by phase error detector **50**.

[0082] Summers **82**, **84** multipliers **86**, **88**, **90**, **91** and delay blocks **92**, **94** are arranged to calculate a loop filter gain k1adapt(n) for each symbol in accordance with the equations,

$$k1adapt(n) = k1adapt(n-1) + 2\gamma\alpha(n-1)error\,\mathrm{Im}ag(n-1))$$
$$\alpha(n) = (1 - k1adapt(n-1) * error\,\mathrm{Re}al(n) * \alpha(n-1) - error\,\mathrm{Im}ag(n))$$

where γ is a small positive auxiliary step size.

[0083] Low pass filter **56** additionally calculates k1adaptavg(n), a filtered, averaged version of k1adapt(n). The structure of an example low pass filter **56** is illustrated schematically in **FIG. 8**. Example low pass filter **56** sums a delayed version of k1adapt, k1adapt(n-1) to the present k1adapt(n) using summer **96** and delay blocks **97**, **98** and multipliers **95**, **99** to calculate

$$k1adaptavg(n) = \lambda*k1adapt(n) + (1-\lambda)*k1adaptavg(n-1)$$

k1adaptavg(n) may be used in the second adaptive mode of filter **52** instead of k1adapt(n). λ controls the bandwidth of filter **52**. Typically λ is selected close to 1. Suitable values of λ and γ may be hard coded, or provided by a controller (not shown) controlling overall operation of receiver **10**.

[0084] Conveniently, adapting the gain and bandwidth of loop filter **52** with the estimate of the phase and frequency error allows carrier loop **30** to more quickly track changes in the phase error, while reducing jitter.

[0085] Advantageously use of a first order loop **52** with a low pass filter, allows loop gain, k1adapt, to be computed by block **54** as a function of phase detector error to minimize error in the decoded symbol y(n) in much the same way as the phase offset estimate $\phi(n)$ is calculated to minimize error in the decoded symbol y(n) by block **52**. This allows adjustment of the bandwidth of loop filter **52** continuously over time. Moreover, as the loop gain k1adapt is adaptive changing on a symbol by symbol basis, loop **52** provides performance of a second order loop which allows tracking of frequency variation, using only a first order loop. Furthermore, this approach takes into account any hidden parametric model affecting the variations of the RF signal frequency. Indeed, if we assume that the phase information at time n can be described as a perturbation of the phase at time n-1, (i.e. that there is a correlation between two consecutives samples in the form described above, namely $\theta(n) = \theta(n-1) + k1\,\Delta$), finding the gain k1 at each time n may improve the estimation of $\theta(n)$.

[0086] Conveniently, receiver **10**, using adaptive loop filter **52** may better track time varying channels such as an RF channel for a moving receiver. Similarly, the example receiver may better track a signal whose frequency is changing over time due to a Doppler shift. This may occur when a vehicle, such as a car or an airplane, is moving in the vicinity of a stationary receiver. Without adapting loop filter **52**, it is difficult to first detect a Doppler shift and then compute optimum gains k**1**, k**2** as these change continuously. If not processed adequately the phase correction would generally lead to a corruption of the symbol y(n) which will lead ultimately to a failure of the equalizer in the tracking mode.

[0087] The above noted description and **FIGS. 1-8** detail a VSB receiver. A person of ordinary skill should now however appreciate that the invention may easily be embodied in other receivers, such as QAM receivers. To that end, **FIG. 9** illustrates a QAM receiver **10'**, similar to VSB receiver **10** and exemplary of an embodiment of the present invention. As noted, however, for QAM, two orthogonal data streams are transmitted at the same time. One constitutes the real portion of a complex data symbol, the other the imaginary

portion. These are typically represented as an I data stream and a Q data stream. For QAM**64**, each symbol's I value takes one of eight relative values: **–7**, **-5**, **-3**, **-1**, **1**, **3**, **5**, and **7**, so does the Q value. For QAM**256** each symbol's I value takes one of sixteen values: **–15**, **-13**, **-11**, **-9**, **–7**, **-5**, **-3**, **-1**, **1**, **3**, **5**, **7**, **9**, **11**, and **15** so does the Q value.

[0088] Receiver **10'** is similar to receiver **10** sharing many identical components. As such, like components have been marked with like numerals and will not be further detailed herein. Components similar to those of receiver **10**, but adapted to operate on a QAM stream have been but including a prime (') symbol. Adaptation of these components to process QAM signals will be apparent to those of ordination skill. For example, slicer **38'** may produce complex signals S(n) in the QAM constellation from complex signals y(n). FFE **26'** and DFE **36'** are formed with complex filter parameter suitable for QAM signals. Interconnection between components reflects complex quantities, including real and imaginary data, transferred between functional blocks.

[0089] Moreover, as the signal is complex, block **32** (**FIG. 1**) is not required, and the phase error may be derived by loop filter **52** from S(n) and y(n). Indeed, use of derot(n) to produce a phase error estimate of the carrier is no longer required. Similarly, phase errors may be calculated by the equivalents of blocks **62a-62e** and **64a-64b** and **100a-100b** using complex valued S(n), y(n) and qpsk (i.e. actual $S_i(n)$, $S_q(n)$, $y_i(n)$, $y_q(n)$ $qpsk_i(n)$, $qpsk_q(n)$) using these blocks. Loop filter, adaptive gain circuit, and low pass filter of carrier loop **30'** may otherwise be formed in manners identical to loop filter **52**, adaptive gain control **54** and low pass filter **56** of receiver **10**.

[0090] As the signal provided to slicer **38'** is complex, carrier recovery circuit **29'** (and particularly mixer **28'**) is downstream of summer 34 (unlike in receiver **10**). Similarly, a further mixer **31'** re-rotates the output of slicer **38'** by an angle of $-\phi(n)$ so as not to lose information in the signal output by slicer **38'**, indicative of a phase error in the received signal.

[0091] Conveniently, as receiver **10** and receiver **10'** use common and

similar components, these receivers could be combined to form a multi-mode receiver capable of receiving VSB or QAM signal. VSB/QAM mode could be toggled externally.

[0092] Of course, the above described embodiments are intended to be illustrative only and in no way limiting. The described embodiments of carrying out the invention, are susceptible to many modifications of form, arrangement of parts, details and order of operation. The invention, rather, is intended to encompass all such modification within its scope, as defined by the claims.